

**ANALOGUE SELF-CALIBRATION METHOD AND APPARATUS FOR LOW
NOISE, FAST AND WIDE-LOCKING RANGE PHASE LOCKED LOOP**

BACKGROUND OF THE INVENTION

Field of the Invention

- 5 The present invention relates to a Phase Locked Loop (PLL) circuit having a variable output frequency. More particularly, the present invention consists of an improved method for a fast and automatic setting of the output frequency of PLL synthesizers that overcomes the classical problems associated with prior implementation, concerning: linearity, locking range limitation, settling time, jitter, phase noise, and
- 10 spectral purity deterioration.

Description of the Prior Art

- A Phase-Locked Loop (PLL) synthesizer is a circuit used to generate a periodic signal with a precise frequency. This kind of circuits is widely used in many communication and measurement products. Its application includes also advanced
- 15 digital systems, such as microprocessors and micro-controllers. The PLL circuits should be designed with stringent constraints in term of noise performances, settling time, power consumption, locking range, integration, cost, etc.

- Typically PLLs include a phase detector that compares the phase of the reference signal to the phase of an internal feedback signal, a charge pump and a low pass
- 20 loop filter for setting an analogue voltage proportional to the detected phase difference, a Voltage-Controlled Oscillator (VCO) that generate a periodic output signal with a frequency proportional to its input voltage, and a frequency divider that generates the feedback signal after dividing the frequency of the output signal by a predefined integer or fractional number (N).

- 25 Two seemingly contradictory requirements constitute the fundamental forces driving the design of VCOs: On one hand, a wide frequency tuning range and thus a high VCO gain are needed to compensate temperature and process variation and to cover the frequency band of the considered application. At the same time, the gain of the VCO should be as small as possible to meet phase noise and spectral purity
- 30 specifications. In fact, the more VCO has a high gain the more its sensitivity to the

noise in its control path increases. The VCO control path is in general affected by several noise sources that include: charge pump noise, filter noise and ripple due to mismatching between the charge pump up and down currents. The noise in the signal path generated by the active elements of the VCO is also dramatically amplified and converted to the phase noise if the VCO gain is high.

Some approaches for simultaneously reducing the VCO gain and enabling a wide frequency range were recently presented as for example in US5942949, in US 6574288 B1 and US2003 0119467 A1 patent documents. All these solutions consist in breaking the wide range tuning curve into several narrower-range sections with some frequency overlap. A digital calibration of a switched-capacitor network is used to choose the appropriate narrow-range section before starting an analogue fine frequency tuning over this curve.

Additional blocks are needed to build a second loop, which achieves the digital self-calibration of the VCO. The second loop is generally constituted by a phase/frequency detector, a charge pump followed by a capacitor or a digital accumulator, and a state machine that applies a sequence of digital control inputs values to the VCO. The complexity of those extra blocks increases proportionally to the needed accuracy of the output frequency. Moreover, since the consecutive digital control words (e.g. 11110 and 00001) can use totally different units of the switched-capacitor network, the accumulated capacitance errors due to process variation can reach very high values. Therefore, a very high frequency overlap is required to compensate these errors, leading to a higher gain of the VCO, a lower total tuning range, and a higher parasitic capacitance. This is why digital calibration can result in prolonged design cycles with a significant additional area and cost of the system.

This digital self-calibration is in general implemented with a simple algorithm such as sequential search as well as with more complex ones such as binary search. However, the time needed for those searching algorithms to achieve a fine resolution is often so high that only a coarse tuning or a self-calibration during the power up of the system is possible. Many side effects can affect the PLL output frequency during its on-mode such as temperature variation, power supply fluctuation, injection pulling etc. Therefore if the frequency varies significantly during the on-mode of the system, the PLL will not be able to correct this variation within a reasonable time. Moreover,

The PLL settling time is one of the most important criteria for many applications. It is even the most critical figure for systems dedicated to fast frequency-hopped spread-spectrum, Ultra-Wideband, data recovery or time-multiplexed transceivers. Therefore, already at the top-level system design it would be desirable to avoid the classical scheme, where a digital calibration for the first coarse tuning of the frequency is used, followed by a continuous tuning for the final fine adjustment of the frequency. This classical scheme constitutes the basis of the methods described in the patent documents US 6574288 B1 and US20030119467 A1. For the fine continuous tuning a reverse-biased diode, nMOS or pMOS varactor are typically used. However an integrated circuit is disclosed in US 6574288 B1 and US20030119467 A1 enabling to avoid the use of these traditional varactors during the fine-tuning stage. In this circuit, capacitors with fixed values are used in an implementation where each capacitor is linked to a variable impedance device. The variable impedance devices are transistors having their gate controlled by a plurality of analogue signals. The equivalent capacitance of the circuit will be indirectly varied by varying the impedance the variable impedance devices. To avoid the abrupt nonlinear variation of the impedances an offset voltage is introduced between the control analogue signals. However, if a fast and fine resolution self-calibration technique is available, it will enable to be very close to the desired frequency. In this way the use of the noisy and cumbersome variable impedance devices, capacitors and their control circuitry can be avoided, and only a small varactor (pn junction, nMOS or pMOS) will be sufficient to achieve the final fine-tuning. Moreover, the gain of the VCO will be at its minimum during the operation of the system and the noise performances will be significantly improved.

SUMMARY OF THE INVENTION

The aim of the present invention is to provide a method and apparatus for a fast and automatic setting of the phase locked loop (PLL) output frequency over a wide tuning range, with a very low VCO gain during the in-lock state to improve the noise performances, and without using the costly, complex and time consuming digital self-calibration techniques.

This aim is achieved with a method for analogue self calibrating of a phase locked loop circuit comprising a phase frequency detector, a charge pump, a loop filter), a

voltage controlled oscillator, including a plurality of VCO tuned elements, which output signal is compared with a reference signal frequency entering in the phase frequency detector characterized in that, the voltage controlled oscillator operating mode, using a linearized frequency versus voltage curve, is switched, in a first
5 frequency tuning operation enabling a wide locking range, to a linear high gain mode, after locking to the appropriate frequency with the said first tuning operation, said voltage controlled oscillator operating mode is automatically switched to a zero-gain mode while keeping the frequency of said voltage controlled oscillator unchanged.

The sensitivity of the said voltage controlled oscillator to the noise in the control path
10 is then practically eliminated and the phase noise performances significantly improved, and optionally, after said zero-gain mode, said voltage controlled oscillator operating mode is switched to a low gain mode enabling a fine tuning of the frequency by the phase locked loop for compensating small residual frequency errors and temperature variations.

15 The present invention concerns also an integrated circuit comprising a phase locked loop circuit including, a phase or frequency detector that compares the phase and frequency of a reference signal to the phase and frequency of an internal feedback signal and generates output error signals, a charge pump that generates amounts of charges proportional to said output error signals, a loop filter for setting an analogue
20 voltage proportional to the charges accumulated in their capacitors, a voltage-controlled oscillator with multiples inputs corresponding each to a VCO tuned element characterized in that said phase locked loop circuit includes a gain mode switcher circuit connected between the loop filter output and the voltage-controlled oscillator inputs, enabling the voltage-controlled oscillator to work successively in a
25 linear high-gain mode and a zero-gain mode; the gain mode switcher includes offsets generators circuits, a switch configuration and comparators; the offsets generators generate the voltages after shifting the loop filter output voltage with predefined offsets; the switch configuration apply the voltages of the offset generators to the inputs of the voltage-controlled oscillator during the linear high gain mode, isolates
30 the inputs of the voltage-controlled oscillator from the offset generators and apply the output voltages of said offsets generators to the inputs of the comparators during the transition to zero gain mode, apply the resulting outputs voltages of said comparators

to the inputs of the voltage-controlled oscillator, and finally freeze the state of each comparator and thus the frequency of the voltage-controlled oscillator making it independent on the loop filter output voltage, which constitute the zero-gain mode.

5 A VCO tuned element corresponds here either to a varactor (voltage controlled capacitor), to a voltage controlled current sources VCI or to any other voltage controlled component varying the frequency of the VCO. A set of these elements respectively constitutes a voltage-controlled oscillator VCO provided with several inputs as described below.

10 In general the present invention is a PLL-based frequency synthesizer having a reconfigurable VCO with three modes of operation: a Linear-High-gain, Zero-gain, and Low-gain mode. During a first tuning operation, a Gain-Mode Switcher sets the VCO in a high gain mode, enabling wide-locking range of the synthesizer with a fast settling time. During this operation, the control voltage of the VCO is varied by the PLL until the appropriate output frequency is found.

15 When the loop is locked, the VCO is automatically switched to Zero-gain mode by the Gain-Mode Switcher, while keeping the frequency of its output signal practically constant. Its sensitivity to the noise in the control path is then practically eliminated and its phase noise performances significantly improved. If the frequency error is sufficiently small for the considered application, the tuning is stopped at this stage. If
20 the error is not sufficiently small the VCO is switched again to Low-gain mode and fine-tuning adjustment of the output frequency is achieved.

Another embodiment of the present invention is a method and apparatus for providing a linear variation of the frequency over all the voltage tuning range. The application of this method is particularly important during the High-gain mode, where
25 the VCO gain variation may leads to serious degradation of the loop stability margin and the noise performances.

The present invention is not limited to charge-pump PLLs. As long as the circuit has a tuning system using a variable capacitance or a variable current, the reconfigurable three-mode (Linear-High-gain, Zero-gain, and eventually Low-gain) varactor (voltage
30 controlled capacitor) or controlled current source using a Gain-Mode Switcher in

accordance with the present invention, can be used for setting the desired value of the capacitance or the current.

BRIEF DESCRIPTION OF THE DRAWINGS

It is noted that the appended drawings illustrate only exemplary embodiments of the invention and are, therefore, not considered limiting of its scope.

FIG. 1 illustrates a block diagram of a charge-pump phase locked loop system in accordance with the present invention;

FIG. 2 diagrammatically shows the Gain-Mode Switcher incorporated into the PLL of FIG. 1;

FIG. 3 diagrammatically shows an exemplary implementation of an offset-generator unit (e.g. 201) of the gain-mode switcher of FIG. 2;

FIG. 4 diagrammatically shows an exemplary implementation of a switcher-comparator unit (e.g. 221) of the gain-switcher of FIG. 2;

FIG. 5 illustrates an exemplary embodiment of a multi-inputs-VCO and their controlled elements in accordance with the present invention and which can be incorporated in the PLL of FIG. 1;

FIG. 6 illustrates another exemplary embodiment of a multi-inputs-VCO and their controlled elements in accordance with the present invention and which can be incorporated in the PLL of FIG. 1;

FIG. 7 shows a hypothetical frequency versus voltage characteristic needed for the considered application;

FIG. 8 the curve 820 shows a hypothetical capacitance versus voltage characteristic needed for the considered application, and C01 to C19 curves show the capacitance versus voltage curves used to build the curve. 1;

FIG. 9 curves V01 to V19 illustrate the output voltages of the offset generators implemented in the gain-mode switcher of FIG. 2. These output voltages are applied to the capacitors units 501 to 519 of the multi-inputs-VCO of FIG. 5;

FIG. 10 illustrates the input 1003 and output 1002 voltage of the comparator 405 illustrated in FIG. 4;

FIG. 11 illustrates the capacitance versus control voltage characteristics of the VCO, during linear-high-gain mode curve 820, zero-gain mode curve 1101, and low-gain mode curve 1102;

FIG. 12 diagrammatically shows the pattern of some internal signals of the PLL system of FIG. 1, and exemplary variations of control voltage (1201, 1202, 1203) during the loop settlings.

DETAILED DESCRIPTION

FIG.1 is a functional block diagram of the charge pump phase locked loop (PLL) system in accordance with the present invention. The PLL circuit, generally designed 100, includes a Phase/Frequency detector (PFD) 121 to compares the phase and the frequency of the reference signal (F_{ref}) to the phase and the frequency of the feedback signal 128, and generates an error signal. The error signal is either an up signal or a down signal depending on the sign of the detected error.

Charge pump CP 122 generates an amount of charge equivalent to the error signal provided by the Phase/Frequency detector PFD 121. Depending upon the polarity of the signal (up or down), the charge is either added to or subtracted from a capacitor in the Low Pass Filter (LPF) 123. Accordingly, the low pass loop filter 123 generates a voltage 124 that will be applied to the input of the Gain-Mode Switcher 125 (GMS).

The Gain-Mode Switcher (GMS) 125 is the block of the PLL 100 that enables to change the gain mode of the VCO 126 during the tuning operations. Three operation modes are possible: A Linear-High-Gain mode (LHG-mode), Zero-Gain mode (ZG-mode), and Low-Gain mode (LG-mode). The Gain-Mode Switcher (GMS) 125 is composed on N parallel paths. N is equal to 19 in the exemplary Gain-Mode Switcher embodiment of FIG. 2. Each path includes an Offsets-Generator OG (e.g., 201) and a Switcher-Comparator SC (e.g., 221).

Each Offset-Generator OG (e.g., 201) receives the voltage V_i 124 at its input and generates an output voltage with an added offset, that is $V_i + \Delta V$ (e.g., 243). The values of the generated offsets will be varied from one Offset-Generator OG to the

other by varying the dimensions of their components. FIG. 3 illustrates an exemplary embodiment of the Offset-Generator OG incorporated in Gain-Mode Switcher (GMS) 125. The dimensions $(W/L)_1$ and $(W/L)_2$ of transistors 301 and 302 are chosen in such way that $(W/L)_2$ is equal to $k \times (W/L)_1$. By changing the value of k , one can
5 change the value of the offset ΔV . The parameter k will be chosen either higher than 1 or lower than 1. The Offset-Generators OG with k higher than 1 will generate a positive offset and those with k lower than 1 will generate a negative offset. The output voltage of each Offset-Generator OG (e.g., 201) is applied to a Switcher-Comparator SC (e.g., 221).

10 Fig. 4 shows an exemplary embodiment of a Switcher-Comparator SC unit of the Gain-Mode Switcher (GMS) 125. In the first tuning, the PLL is started in LHG-mode, each Switcher-Comparator SC (e.g., 221) of the Gain-Mode Switcher (GMS) 125 will enable to apply its input voltage $V_t + \Delta V$ (e.g., 243) directly to one of the inputs of the VCO (e.g., 101). After a first locking, the input voltage $V_t + \Delta V$ of each Switcher-
15 Comparator SC will be compared to V_{ref} 406, and its output switched to zero or to the power supply voltage V_{cc} depending upon the $V_t + \Delta V$ is higher or lower than V_{ref} . In this example, V_{ref} corresponds to the middle of the capacitance versus voltage characteristic of the used varactors (among 501 to 519). The capacitance of these varactors is thus switched to its maximum C_{max} or to its minimum C_{min} ,
20 depending upon its value was at C_{max} or at C_{min} when the PLL was locked at the end of the LHG-mode. By this operation the mode is switched to ZG-mode while keeping the total capacitance and so the output frequency unchanged.

The present invention can be applied on two types of PLL configuration:

a) The PLL with a capacitance controlled VCO, such as Inductance – capacitance
25 LC oscillator, which is driven by a voltage varying a set of capacitances (varactors). This PLL is often used in radio-frequency (RF) applications for large frequency ranges. The main advantage of LC VCO is its low phase noise.

FIG. 5 shows an exemplary of the multi-inputs VCO incorporated in FIG. 1. This schematic is an embodiment of a modified inductance L - capacitance C VCO (LC
30 VCO), which has several variable capacitors (e.g., 501 to 520 in FIG. 5). The values of these capacitors units 501 to 520 are controlled by input voltages 101-120

supplied by the Gain-Mode Switcher (GMS) 125. The output frequency of the VCO depends on both the value of the inductance 521 and the value of the capacitors 501 to 520. The active circuit AC 522 generates the required negative impedance for the circuit to oscillate.

- 5 b) The ring oscillator PLL, which is driven by a voltage varying a set of current sources. This PLL is preferably used in digital applications and/or at low frequency ranges. The main advantage is this type of PLL is its compactness and wide tuning range performance.

FIG. 6 shows an example of VCO using voltage controlled current sources (VCI) that
10 can be incorporated in FIG. 1 to replace the LC VCO as shown in FIG. 5. The voltage controlled oscillator VCO is made by a current controlled oscillator ICO preceded by voltage - current V-I converters at each input. The values of the currents 601 to 620 generated by the converters are controlled by input voltages 101-120 supplied by the Gain Mode Switcher (GMS) 125. The output frequency of the oscillator depends then
15 on the values of the currents 601 to 620. The ring oscillator 621 can be implemented with a cascade of differential buffer delay stages as well as with an odd number of inverters.

The present invention will be clarified with the help of the hypothetical application illustrated in FIG. 7. PLL's are often designed for a wide range of frequencies. The
20 wide frequency tuning range is needed to compensate temperature and process variation and to cover the frequency band of the considered application. In the case of the application illustrated in Fig. 7, a frequency tuning range of 800 MHz is needed. The PLL should be able to set any frequency in the band 1 GHz to 1.8 GHz. The available tuning voltage is about 2 V (i.e., from 0.5 V to 2.5 V), thus, the needed
25 VCO gain (K_{VCO}) is 400 MHz/V. It is well known that the operating curves (i.e., input voltage versus output frequency) of classical VCOs are in general highly non-linear. A variation of the gain K_{VCO} will change the open-loop gain and hence change the loop bandwidth and phase margin. This is why in this hypothetical application of FIG. 7 a linear operating curve with a constant K_{VCO} is targeted.

30 The curve 820 of FIG.8 illustrates the needed capacitance versus voltage characteristic of the VCO varactor to produce the operating curve 701. 820 is

deduced from 701 by using the well-known formula $2\pi F = (LC)^{-1/2}$, where F is the frequency of the VCO output signal, and L and C are respectively the equivalent parallel inductance (521) and capacitance (501-520) of the tank. To simplify the analysis, the fixed capacitance of the VCO is supposed to be negligible and its inductance 521 constant and equal to 2 nH. To built the wide-range curve 820, we start by breaking it into several narrower-range sections, 801 to 819. Each narrower-range section (e.g, 801) will be represented by an individual curve (e.g., C_{01}) having the same variation over the corresponding voltage interval (e.g., 0.1V in FIG. 8). As result, the curve 820 can be reconstituted entirely by a simple addition of the individual curves C_{01} to C_{19} . Each one of the curves C_{01} to C_{19} will correspond to a given varactor among the varactors 501 to 519 of FIG. 5.

The exemplary varactors 501 to 519 are MOS (Metal Oxide Semi-conductor) transistors, dimensioned in such way that there capacitance characteristics has respectively the same variation of C_{01} to C_{19} . In the exemplary embodiments of FIG. 5 the control voltage 124 is applied to the MOS capacitors 515 and generates the operating curve C_{15} . To generate the other operating curves $C_{15\pm i}$, that is C_{01} to C_{15} curves, a tuning voltage with a various offsets are applied to the corresponding MOS capacitors. More precisely, to generate the curve $C_{15\pm i}$, a voltage $V_{15\pm i} = V_t \pm 0.1 \times i$ will be applied to the corresponding MOS capacitor. This voltage is one among the voltages V_{01} to V_{19} illustrated in FIG. 9, which illustrate the output voltages of the offset generators implemented in the gain-mode switcher Gain-Mode Switcher (GMS) 125 of FIG. 2. Therefore when those output voltages are directly applied to the MOS capacitors 501 to 519 of the multi-inputs VCO of FIG. 5, the PLL works in the LHG-mode and the resulting frequency versus tuning voltage curve is exactly the desired curve 701 of FIG. 7.

FIG. 12 diagrammatically shows the pattern of some internal signals of the PLL system 100, and exemplary variations of control voltage during the three modes of operations.

During a first tuning operation, the Gain-Mode Switcher (GMS) 125 sets the Multi-Inputs VCO 126 in the LHG-mode 1201. During this operation, switch 240 is open and 241 closed. Accordingly, the constant voltage V_{∞} 220, is provided to varactor 520, freezing its capacitance in the middle of its capacitance versus voltage

characteristic. On the other hand, switch 401 of each Switcher-Comparator SC (among 221-239) of the Gain-Mode Switcher (GMS) is closed while 402, 403, 404 are open. Accordingly, the outputs voltages of the Offsets-Generators OG 201-219 are directly provided to varactors 501 to 519. The phase frequency detector PFD 121 and the charge pump CP 122 will vary the control voltage 124 and thus, through the Offset-Generators OG 201-219, will also vary the voltages 201-219 according to the curves $V_{01}-V_{19}$ of Fig.9. During this tuning operation, the frequency is varied according to the wide-range curve 701 until the appropriate value is found. In the given example V_t 124 will be set to 1.5V (702) and output frequency to 1.42 GHz (703). The noise performance of the PLL is not critical at this stage. Therefore, in this mode the PLL can be optimized to be stable and have a fast settling time even if it will be at the cost of more noise. For this, a fraction of the filter capacitances can be optionally switched off and/or the current of the charge pump increased.

The duration of the LHG-mode 1201 is controlled by a Lock Detector LD 129. This mode is selected during a sufficiently long time for the loop to lock. When the control voltage 124 will converge to the needed value (e.g., 1.5 V 702), the Lock Detector LD send a signal to the Gain-Mode Switcher (GMS) and the VCO is switched to Zero-gain mode (ZG-mode in FIG. 12) by the Gain-Mode Switcher (GMS) 125, while keeping its frequency constant.

During the ZG-mode 1202 step, the switch 401 of each Switcher-Comparator SC among 221 to 239 is open, isolating the varactors 501-519 from the output voltages $V_{01}-V_{19}$ of Offset-Generators (OG) 201-219. At the same time switch 403 and 402 are closed. Voltages $V_{01}-V_{19}$ will thus be compared to V_{ref} 406 by the comparator 405. V_{ref} 406, is chosen to correspond to the middle of the capacitance versus voltage characteristic C15 of FIG. 8, which corresponds to varactor 515 (e.g., $V_{ref} \gg 0.95$ V). The output of 405 will then be switched to zero or to V_{cc} depending upon the input voltage is higher or lower than V_{ref} . Accordingly, the capacitance of the varactors 501-519 will be switched to its maximum value or to its minimum value. To freeze the situation and to make it independent on future variations of the control voltage 124, switch 402 is opened and 404 is closed. The capacitance of the tank 501-519 are thus frozen and correspond to a constant capacitance (e.g., operating curve 1101), which is independent on V_t 124. The sensitivity to the noise in the control path is

then practically eliminated and the phase noise performances of the PLL 100, significantly improved. At this stage, the equivalent output frequency of the VCO is equal to the desired frequency (e.g., 1.42 GHz 703) with a small error. The error can be reduced by increasing the number of fractions in FIG. 8 and thus increasing the number of VCO varactors in FIG. 5. If the frequency error is sufficiently small for the considered application the tuning is stopped.

If the error is not sufficiently small the VCO is switched to Low-Gain (LG) mode 1203. Switch 241 is opened and switch 240 is closed. The tuning voltage 124 is then applied to an additional varactor 520, and fine-tuning adjustment over operating curve 1102 is achieved.

Optionally, comparators are added to the Gain-Mode Switcher (GMS) 125 to set an upper and a lower limit for its input signal 124. Accordingly, these comparators will control the value of the signal 124. If during LG-mode 1203, this signal reaches either of the upper limit 1104 or the lower limit 1103, a Gain-Mode Switcher (GMS) 125 will be restarted, and LHG-mode 1201 is selected again. The limits are preferably set near Vcc for the upper (1104) and near zero for the lower (1103). This operation is useful if the fine-tuning range is not sufficiently wide to compensate frequency variation that can arise during LG-mode 1203. This frequency variation can be caused by a rise in the chip temperature for instance.

To preserve the stability of the PLL in both LHG-mode and LG-mode, the product of the charge pump CP 122 current and the gain of the VCO 126 will be kept constant. Therefore, a fraction of the charge pump CP current will be switched off during LHG-mode 1201 and switched on during LG-mode 1203. Optionally, a fraction of the filter capacitance is switched off during LHG-mode 1201, resulting in significant improvement of the PLL settling time at the cost of a worst noise performance. In fact, the noise performance becomes critical only in the final LG-mode.

Another option particularly suitable for low voltage applications is to add a voltage doubler circuit that increase the voltage supply of the charge pump (CP), the loop filter and the offsets generators (OG) during the linear high-gain mode (LHG) and hence enhance the tuning range of the PLL. A switch configuration enabling the isolation of the noisy voltage doubler during the low gain (LG) mode and the

application instead, the low noise output operating voltage supply is optionally added to save PLL spectrum purity.

The forgoing description has been presented for the purposes of illustration and description. It is not intended to be exhaustive or to limit the invention to the
5 exemplary embodiments disclosed, and obviously many modifications and variations are possible in light of the above teaching. For example, the LHG-mode can be implemented with a single input VCO, controlled by a single analog voltage without using any offset generator. The varactors 501 to 519 will be controlled by the same signal. A circuit will be added to change the threshold voltages of the varactors and
10 thus to enable their capacitance to vary exactly as the curves C_{01} - C_{19} of Fig. 8. The switching to the ZG-mode will be achieved by comparing the single control voltage to a plurality of reference voltages and decide for each varactor if it will be set to its maximum value or to its minimum value.

In addition, further modifications and alternative embodiments of this invention will be
15 apparent to those skilled in the art in view of this description. For example, the analogue self-calibrating method which consist on using a linear high gain mode to adjust the frequency and switch to zero gain mode to improve the noise performances, can be used in a successive approximation procedure. The PLL will use successive LHG-mode / ZG-mode stages; in each stage the specified gain of the
20 LHG-mode will be lower than in the precedent stage. By this procedure the tuning will converge successively toward the desired frequency. Therefore, It is to be understood that the forms of the invention herein shown and described are to be taken as exemplary embodiments. Various modifications and alternative embodiments can be used without departing from the scope of the present invention.